



RESEARCH DEPARTMENT



REPORT

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# VSB 2-PSK: A modulation system for digital sound with television

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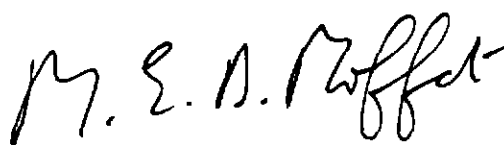
## VSB 2-PSK: A MODULATION SYSTEM FOR DIGITAL SOUND WITH TELEVISION

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### Summary

*This report describes VSB 2-PSK, a modulation system suitable for the transmission of digital sound and data associated with a television signal. This modulation system shows good spectral efficiency and little sensitivity to the type of channel imperfections associated with typical television transmission channels.*

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<b>Section</b>	<b>Title</b>	<b>Page</b>
	<b>Summary.....</b>	<b>Title Page</b>
<b>1.</b>	<b>Introduction .....</b>	<b>1</b>
<b>2.</b>	<b>Description of VSB 2-PSK .....</b>	<b>1</b>
	2.1. Direct approach .....	1
	2.2. Approach through MSK.....	2
	2.3. Division of filtering.....	2
	2.4. Spectrum of transmitted and received signals .....	4
<b>3.</b>	<b>Implementation of the system .....</b>	<b>5</b>
<b>4.</b>	<b>Predicted performance.....</b>	<b>6</b>
	4.1. Noise .....	6
	4.1.1. The effect of noise .....	6
	4.1.2. Expected noise levels.....	7
	4.2. Sensitivity to carrier phase perturbations and sampling timing error .....	8
	4.3. Sensitivity to channel imperfections .....	10
<b>5.</b>	<b>Experimental work .....</b>	<b>11</b>
	5.1. Experimental configuration.....	11
	5.2. Performance in the presence of noise .....	13
	5.3. Effect of carrier phase errors.....	13
	5.4. Effect of multipath propagation.....	13
<b>6.</b>	<b>Conclusion .....</b>	<b>15</b>
<b>7.</b>	<b>Acknowledgements .....</b>	<b>15</b>
<b>8.</b>	<b>References .....</b>	<b>16</b>
	<b>Appendix 1 .....</b>	<b>17</b>
	<b>Appendix 2.....</b>	<b>17</b>

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## 1. Introduction

For many years there has been interest in improving the quality and the number of sound channels associated with television signals. This has resulted in several different proposals for broadcasting two sound channels with terrestrial television,<sup>1,2,3,4</sup> and multiple sound for DBS, for example the C-MAC/packet<sup>5</sup> proposals for satellite broadcasting. There is now a proposal for digital sound transmission with terrestrial television in the UK.

Digital sound transmission has many attractive features. Not only does the basic quality of a digital system compare favourably with the quality of the best analogue transmission systems, but it is also capable of rugged transmission. In analogue systems, the sound quality is often degraded by interference from the associated television picture. In a digital system the effect of this can be minimised. Thus good sound quality can be maintained much more easily over a wider range of reception conditions.

The European Broadcasting Union<sup>6</sup> (EBU) recommended vestigial sideband binary phase shift keying (VSB 2-PSK) for use as a digital modulation system in satellite transmissions using a subcarrier. The intention was to transmit digital sound in association with a PAL or SECAM vision signal. The EBU developed the system to the extent that a specification now exists for a complete package for satellite broadcasting based on this principle. In the discussions it became clear that there were many modulation systems (including QPSK<sup>7</sup>) which may have been suitable. There was little difference between them in their bandwidth requirements or sensitivity to noise. However, VSB 2-PSK was adopted because of its lower sensitivity to errors in the phase of the recovered carrier, i.e. the carrier used for synchronous demodulation.

Previous studies<sup>8</sup> have shown that it is possible to specify a television system with digital sound which is compatible with existing receivers and which provides adequate quality even in poor reception areas. This system is similar to the standard System I (PAL television signals plus an analogue sound carrier 6.0 MHz above the vision). The differences are that the analogue sound carrier is reduced in power level to 10 dB below that of the

vision carrier and a digital signal is added at a frequency above the analogue sound signal. In all the tests so far, quadrature phase shift keying (QPSK)<sup>7</sup> has been used. Experimental equipment was available and so it was convenient to use this in the tests. This modulation system performed well.

As for DBS, VSB 2-PSK may be suitable for use as a digital modulation method for terrestrial television. It occupies a similar bandwidth to QPSK, has similar noise properties but has less sensitivity to carrier phase errors. An important source of additional degradation to the signal under marginal reception conditions can be unwanted incidental phase modulation, introduced at the transmitter, the receiver, or both.

This report describes VSB 2-PSK. It explains its properties in the presence of noise and distortion and describes suitable methods of modulation and demodulation. The theoretical studies draw heavily on the work of the EBU experts group which was set up to consider satellite broadcasting. The only major distortion considered here which was not considered by that group is the effect of echoes (these are not normally a significant degradation in a satellite system).

## 2. Description of VSB 2-PSK

VSB 2-PSK is a modulation system whose signals can be generated by several different techniques. Those not familiar with the range of modulation systems currently being studied are advised to concentrate on the direct approach from 2-PSK. However, those seeking a fuller understanding will wish to consider the approach via minimum shift keying (MSK) (which may also be viewed as a variant of offset-QPSK or alternatively, frequency shift keying).

### 2.1. Direct approach

VSB 2-PSK is a modulation system aptly described by its name. It is a binary phase shift keyed signal (sometimes known as phase reversal keying) whose spectrum occupancy has been reduced by a vestigial sideband filter.

Vestigial sideband transmission is well known in broadcasting. The existing UK terrestrial television standard uses vestigial sideband transmission

for the vision signal. This gives some of the spectrum saving benefits of single sideband transmission without putting too great a requirement on the receiver technology. Although there are some differences between standards in Europe, the overall filtering in a vestigial sideband channel is always anti-symmetric about the vision carrier frequency.

In the case of VSB 2-PSK, the shape of the vestige depends on the application. However, not only must the overall filtering be anti-symmetric about the carrier frequency, but the data must also be filtered in such a way that there is no inter-symbol interference, i.e. the data is filtered anti-symmetrically about a frequency equal to half the bit rate.

The proposal adopted by the EBU is based on transmission of the upper sideband with 50% cosine roll-off (CRO) filtering for both the data and the vestigial sideband i.e. the overall filtering is given by

$$H(f) = \frac{1}{2}(1 + \cos 2\pi(f - (f_c + f_R/4))/f_R);$$

$$f_c - f_R/4 \leq f \leq f_c + 3f_R/4 \quad (1)$$

$$= 0 \text{ otherwise}$$

where  $f_c$  is the carrier frequency and  $f_R$  is a frequency equal to the bit rate. This response is shown in Fig. 1(a) and the resulting power density spectrum is shown in Figs. 1(b) and 1(c).

It is of course possible to adopt different data shaping. Tighter filtering will reduce the bandwidth required, but make carrier recovery and clock recovery harder. The EBU proposal is the solution which is least critical in these respects.

## 2.2. Approach through MSK

VSB 2-PSK can be viewed as a filtered version of minimum shift keying (MSK). This is an important feature because the EBU studies approached VSB 2-PSK as a special case of MSK. Thus some of the details can only be fully understood if this route is followed.

MSK is frequency shift keying with a modulation index of 0.5 and phase continuity between successive data periods. Thus when a logical '1' is transmitted the phase of the carrier advances linearly by  $\pi/2$  during the bit interval; when a logical '0' is transmitted the phase retards linearly by  $\pi/2$  over the bit interval. The modulator can use a frequency modulator, but more often a quadrature phase modulator is employed.

Amoroso and Kivett<sup>9</sup> have shown that it is

possible to generate MSK by a linear filtering of 2-PSK. As a corollary it is possible to demodulate MSK as if it were a 2-PSK signal i.e. using a single mixer. This approach is often called simplified MSK; not because the MSK signals are different – they are identical – but because the modulator and demodulator are simpler. The linear filtering necessary to convert from 2-PSK to MSK is shown in Fig. 2. It can be seen that the filtering is not symmetrical about the carrier frequency. Instead there is significant filtering of, in this case, the lower sideband, giving a signal, MSK, as shown in Fig. 3, which has started to look like VSB 2-PSK.

VSB 2-PSK can be generated from MSK therefore, by additional filtering. The generation of MSK from 2-PSK has gone part way. All that is necessary is to define the appropriate filters to complete the task. This problem is complicated by the need to divide the filtering between transmitter and receiver.

## 2.3. Division of filtering

In any digital system the overall filtering is divided. Some filtering is applied to the signal before transmission and the remainder on reception. The main purpose of filtering applied at the transmitter is to restrict the energy of sidebands which may interfere with other signals. The receiver filter serves to remove interference from other signals and to minimise the noise bandwidth of the receiver.

It is common practice in communication theory to divide the filtering equally between transmitter and receiver. This allows reception using a matched filter: a configuration which is known to have optimum noise performance.

The argument is not so simple in this case. Although equal division of filtering can be proposed, it is not necessarily optimum from the point of view of the receiver manufacturer. The optimum from this viewpoint is that the receiver filter be cheap and non-critical to produce whilst giving results close to the optimum matched-filter configuration.

The EBU studies showed that a satisfactory system could be achieved, based on overall filtering with a 50% cosine roll-off. This could be obtained by starting with MSK, filtering on transmission with a 4-pole Butterworth filter and using a Gaussian receive filter. This work was based on studies of MSK originally performed by Pommier,<sup>10</sup> which indicated that a Gaussian filter was capable of giving very good reception quality without being critical of alignment. This original proposal was fine tuned, and during the course of the studies it was recognised that the resulting transmit filter is wider

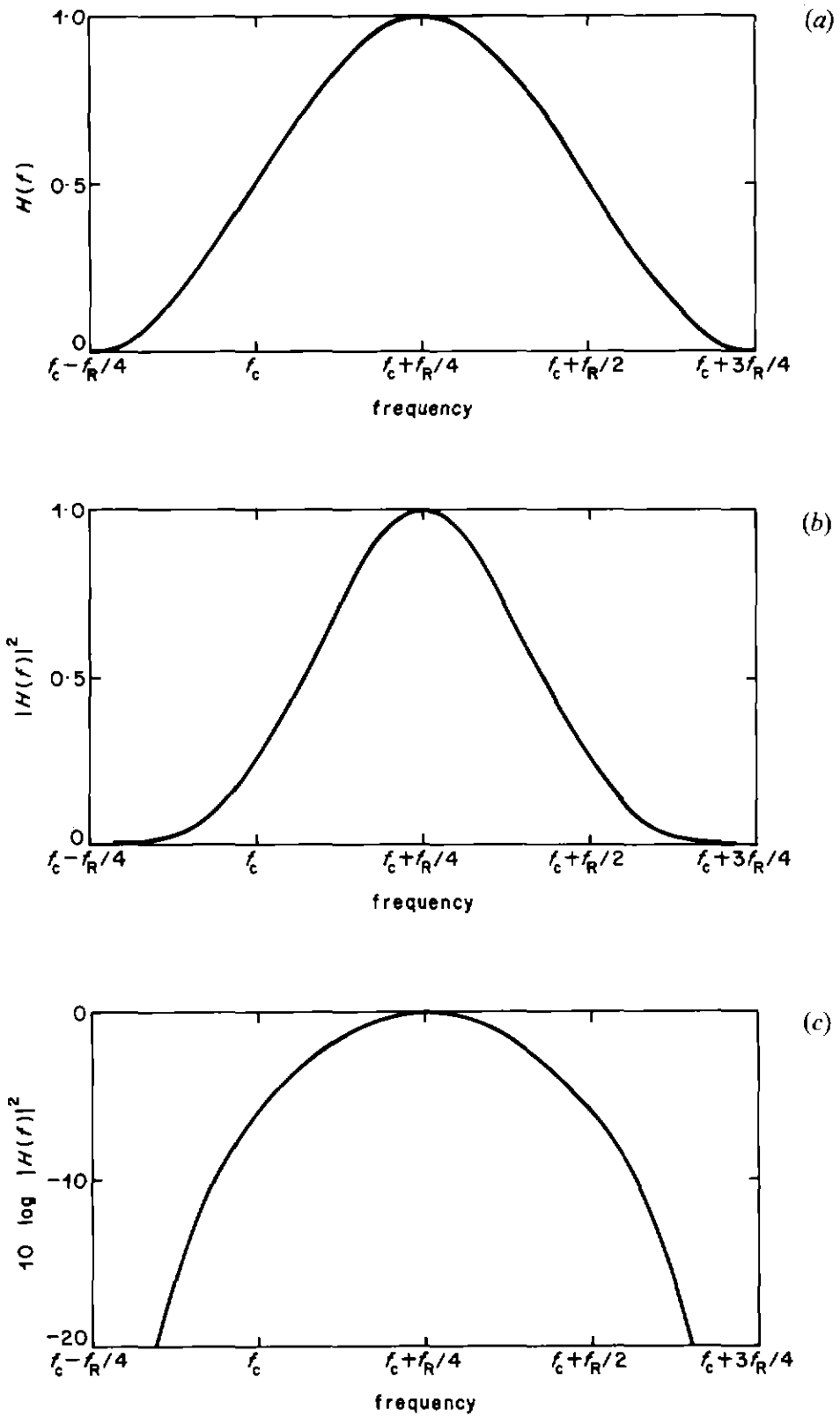


Fig. 1 - VSB 2-PSK filtering (50% CRO), (a)  $H(f)$ , (b)  $|H(f)|^2$ , (c)  $10 \log |H(f)|^2$ .

than that which would normally be used and the filter in the receiver is of narrower bandwidth. This division of filtering is both simple and non critical, and is important in areas where the frequency or bandwidth restricts the use of SAW and ceramic filters.

#### 2.4. Spectrum of transmitted and received signals

It is possible to derive the transmitted and received spectra from the description given in the previous Section.

Let us approach the problem through MSK. It is necessary first to derive the MSK from 2-PSK and then filter it with transmit and receive filters.

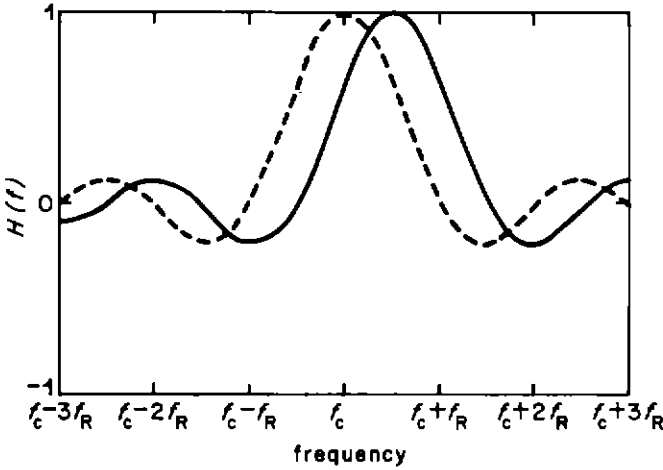


Fig. 2 - Spectrum of 2-PSK and filter response required to generate MSK,

----- 2-PSK,  
—— Filter response needed.

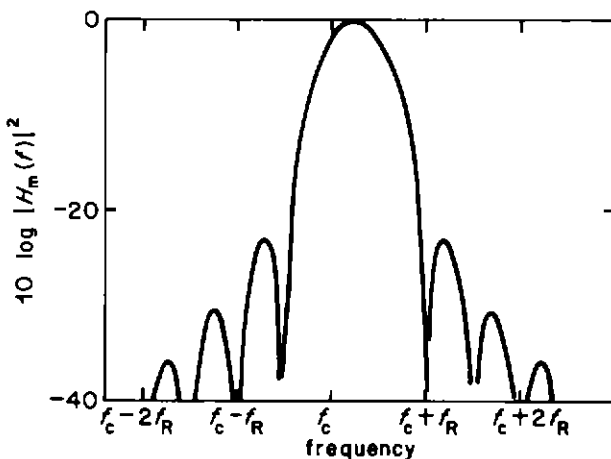


Fig. 3 - Spectrum of MSK.

In the approach through MSK the 2-PSK signal is generated with a box-car modulation, i.e. the signal spectrum is not whitened. The 2-PSK spectrum is therefore

$$H(f) = \frac{\sin \pi(f-f_c)\tau}{\pi(f-f_c)\tau} \quad (2)$$

where  $\tau$  is the bit period ( $1/f_R$ ).

To generate MSK this is filtered by a filter with an impulse response of

$$h(t) = \sin 2\pi(f_c + f_R/2)t; \quad 0 \leq t \leq \tau \quad (3)$$

= 0 otherwise

Provided certain additional conditions are met,<sup>10</sup> MSK is generated with a power spectrum

$$|H_M(f)|^2 = \frac{8}{\pi^2} \cdot \frac{1 + \cos 4\pi v\tau}{(1 - 16\tau^2 v^2)^2} \quad (4)$$

where  $v = f - (f_c + f_R/4)$

The crucial step is now to define additional filtering which will lead to a transformation from MSK to VSB 2-PSK. This step was identified by Pommier and Harris in the course of the EBU studies, and as noted earlier, comprises a cascade of Butterworth and Gaussian filters.

The fourth order Butterworth filter has a bandwidth (between the half power points) of 0.73 times the bit-rate ( $R$ ). Its response is

$$|H_B(f)|^2 = \frac{1}{1 + (2v/0.73R)^8} \quad (5)$$

The Gaussian filter has a bandwidth (between half power points) of 0.45 times bit rate. The response is

$$H_G(f) = \text{alog}_{10}[-0.6(v/0.45R)^2] \quad (6)$$

The overall filtering is thus given by

$$H(f) = H_M(f)H_B(f)H_G(f) \quad (7)$$

The resulting spectrum is shown in Fig. 4. The match between this and the spectrum of Fig. 1(c) is very good.

It is interesting to note that VSB 2-PSK and QPSK occupy the same bandwidth. Indeed VSB 2-PSK with 50% CRO filtering has the same power spectral density as QPSK with 100% CRO filtering. The difference in CRO rates is balanced by the different symbol rates: that of QPSK being half

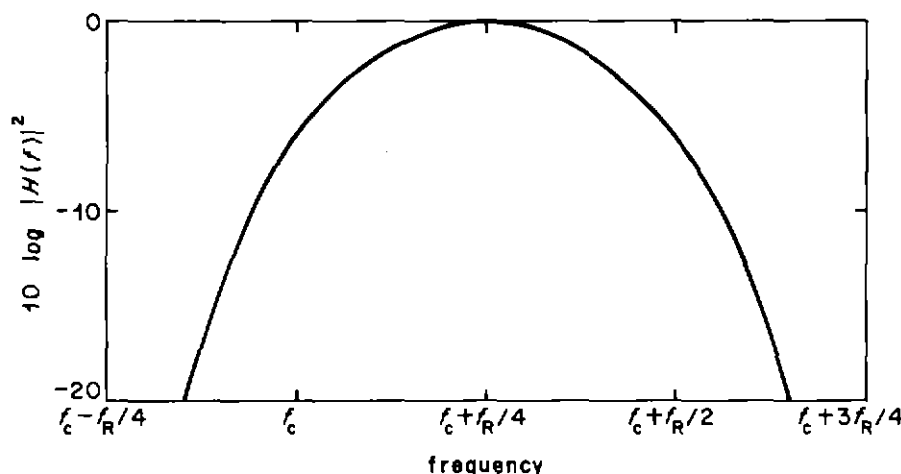


Fig. 4 – Spectrum of VSB 2-PSK generated from MSK.

that of VSB 2-PSK. Thus the spectral densities of the two systems are directly comparable. (Of course if different CRO values had been used then different spectra would result). Although the amplitudes of the spectra are the same, the phases are not. VSB 2-PSK is a signal with near constant envelope. QPSK on the other hand has a large amplitude variation. Fig. 5 shows the i.f. diagram of a VSB 2-PSK signal. This shows the way the instantaneous amplitude and phase vary as a function of time. The diagram shows the locus of the signal vector with the unmodulated carrier as a reference. Fig. 5(a) shows the i.f. diagram of the transmitted signal. Zero carrier would be represented by a point at the centre of the 'circle'. There is little amplitude vari-

ation indicating that VSB 2-PSK can, unlike QPSK, be passed through non-linear systems with little degradation. Fig. 5(b) shows the i.f. diagram of the signal after the filter in the receiver.

### 3. Implementation of the system

The transmission system can be implemented in several ways, each corresponding to one of the approaches noted in Section 2.

Fig. 6 shows the direct generation from 2-PSK. This simply uses a filter to shape 2-PSK. The filter is not simple if built as one filter. However, it can be built in two stages. Firstly, generating MSK from

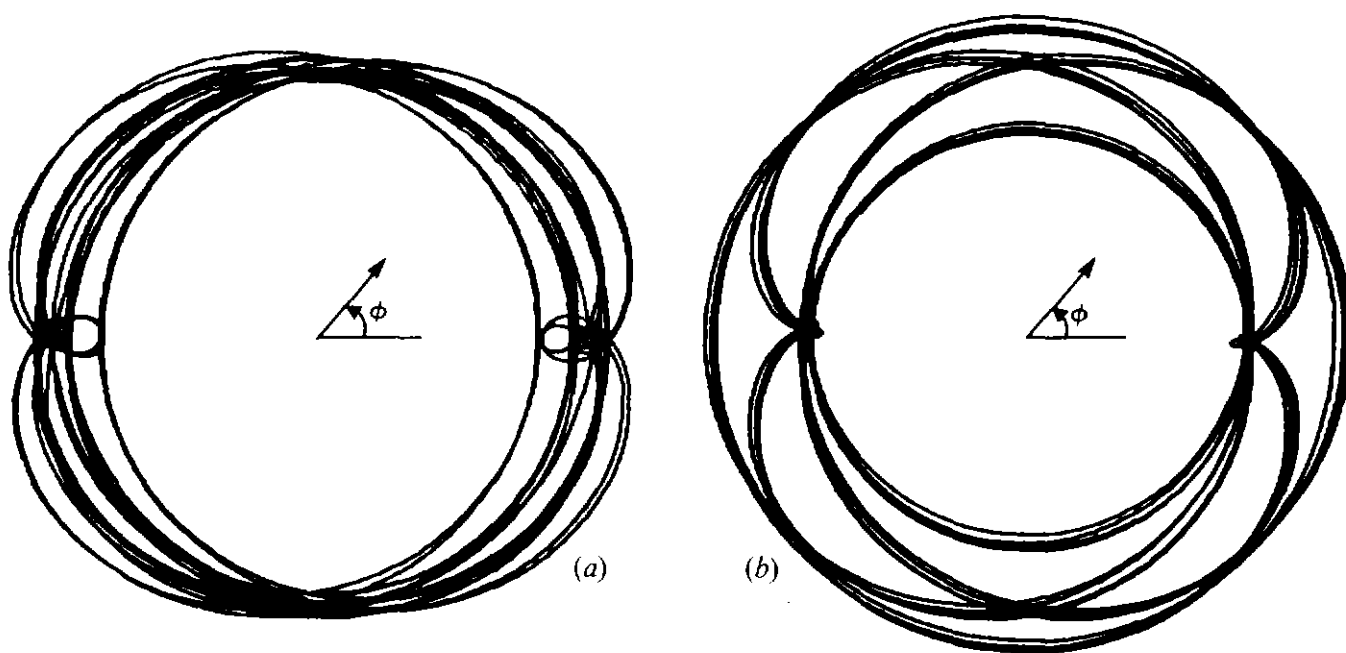


Fig. 5 – I.f. diagrams of VSB 2-PSK,  
(a) As transmitted,  
(b) After the receive filter.

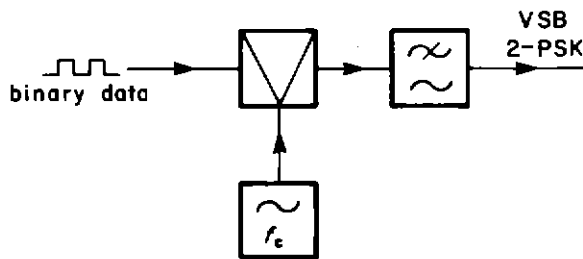


Fig. 6 – Generation of VSB 2-PSK.

2-PSK then including additional filtering as noted in Section 2.4. Alternatively, the filtering could be implemented using computer optimised techniques to obtain the desired shape directly. A SAW filter would be most suitable for this type of optimisation.

On the other hand, the modulator could be based directly on an MSK modulator. This could in turn be based on an FSK system or offset-QPSK, and can be implemented using analogue or digital techniques.

The demodulator can be implemented in several ways. Each method is based on one of the above methods of modulation, although of course it does not matter which method of modulation was in fact used, as the transmitted signals should be the same in each case.

Optimum performance in a noisy channel can be obtained by synchronous demodulation of the received signal.

Fig. 7 shows a typical receiver based on conventional principles for reception of a signal transmitted using a vestigial sideband. The incoming signal is first filtered to ensure correct shaping of the vestige and to ensure a minimum of noise and interference reach the receiver. The carrier is regenerated and used to synchronously demodulate the signal. The regeneration of the carrier can be accomplished by several techniques. The method first proposed was to frequency double the incoming signal. However,

there are other techniques which rely on feedback of an error signal: a modified Costas loop for example, or feedback to maximise the eye-height. At low bit-rates, it is possible to use digital techniques to identify the optimum sampling point and the maximum eye-height, thus providing a useful tool for minimising the effect of distortion (see Section 4.3).

Alternatively, a receiver could use differential (delay-line) demodulators or discriminator demodulation. These demodulators will perform less well than synchronous demodulators and provide no significant advantages, so will not be considered further. Quadrature demodulators can be used, but these introduce extra complexity with no improvement in quality. This option will not be considered further either.

## 4. Predicted performance

### 4.1. Noise

#### 4.1.1. The effect of noise

If the filtering in the system is arranged so that

- the Nyquist restrictions are complied with, and
- the receiver uses a matched filter,

then the probability of receiving a bit erroneously ( $P_e$ ) is given by

$$P_e = \text{erfc}' \sqrt{2E_b/N_0} \quad (8)$$

where  $N_0$  = single sided noise power spectral density incident on the receive filter.

$E_b$  = signal energy per bit incident on the receive filter, i.e. the mean power divided by the bit rate.

$$\text{erfc}'(x) = 1/\sqrt{2\pi} \int_x^\infty e^{-y^2/2} dy$$

$P_e$  is also known as the bit error rate (BER).

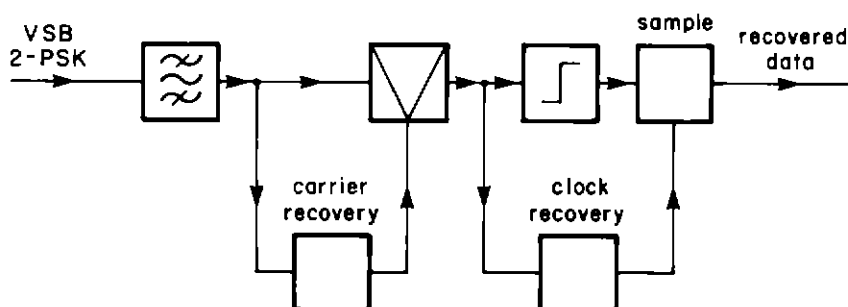


Fig. 7 – Reception of VSB 2-PSK.

A curve of  $P_e$  as a function of the energy density ratio,  $E_b/N_o$ , is shown in Fig. 8.

In the system proposed, the two conditions outlined above are not met. As a consequence the probability of error will be more than indicated by equation 8. However, the degradation predicted by computer simulation is equivalent to an increase in noise level of about 0.1 dB.

#### 4.1.2. Expected noise levels

If it is proposed to use the VSB 2-PSK carrier in conjunction with a television signal, the system designer must ensure that the digital channel (which carries the sound in most proposals) is not the limiting factor in the overall performance of the channel. The video signal occupies greater bandwidth and needs most power, and therefore it is essential that the digital sound does not degrade to unacceptable quality levels before the vision signal.

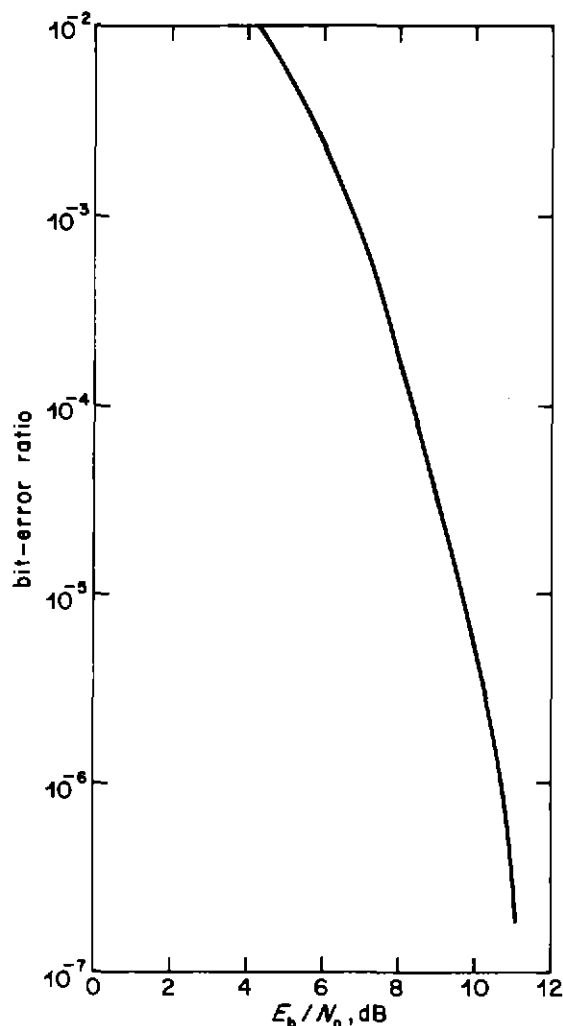


Fig. 8 – Relationship between error rate and noise level.

Two transmission channels have been considered in detail.

(a) *Terrestrial television using vestigial side-band amplitude modulation for the vision signal.* For PAL System I transmissions the VSB 2-PSK signal is added to the conventional signal about 0.55 MHz above the f.m. sound carrier. This frequency is chosen to minimise interference to and from the f.m. sound signal and adjacent channel vision signals. The level of the subcarrier needs to be chosen such that vision failure occurs before sound failure, and such that interference from the new carrier is minimised. These requirements, although pulling in different directions, can be satisfied simultaneously.

Appendix 1 shows that the relationship between video signal-to-noise ratio\* and  $E_b/N_o$  in the digital channel (of 728 kbit/sec, 20 dB below peak-vision power) is given by

$$E_b/N_o = S/N - 3.6 \text{ (dB)} \quad (9)$$

Let us take an extreme example. Consider a viewer receiving a signal with a  $S/N$  of 20 dB. This gives a very poor picture not normally considered to provide an adequate service. If we wish the BER in the digital channel to be about  $10^{-3}$  under the same conditions, a figure which gives significant degradation but at which complete failure has not been reached, we need an  $E_b/N_o$  of about 6.8 dB. Thus from equation 9 we have a margin of about 10 dB. This may well be necessary to protect from the gross attenuation of the digital carrier caused by multi-path propagation, as well as other imperfections introduced by the transmission system.

(b) *Satellite broadcasts using frequency modulation for the multiplexed signal.* VSB 2-PSK was originally considered for a digital subcarrier added to a f.m. television signal for satellite broadcasting. Appendix 2 shows that in a pre-emphasised channel, video signal-to-noise ratio, carrier-to-noise ratio and  $E_b/N_o$  are related by

$$S/N = C/N + 18.5 \text{ (dB)} \quad (10)$$

$$E_b/N_o = C/N + 20 \log_{10} \Delta f / 2.5 - 20 \log_{10} f_o / 7 - 10 \log_{10} R / 2 + 2.3 \text{ (dB)}$$

In practice the lower limit on vision quality is often taken to be vision threshold (i.e.  $C/N \approx 10$  dB). If we make the same assumptions as for terrestrial television about the desirable BER, we find that we have a margin of about 5.5 dB.

\* Note  $E_b/N_o$ ,  $C/N$  and  $S/N$  represent power ratios expressed either in linear terms or dB. The former applies unless dB is explicitly stated.

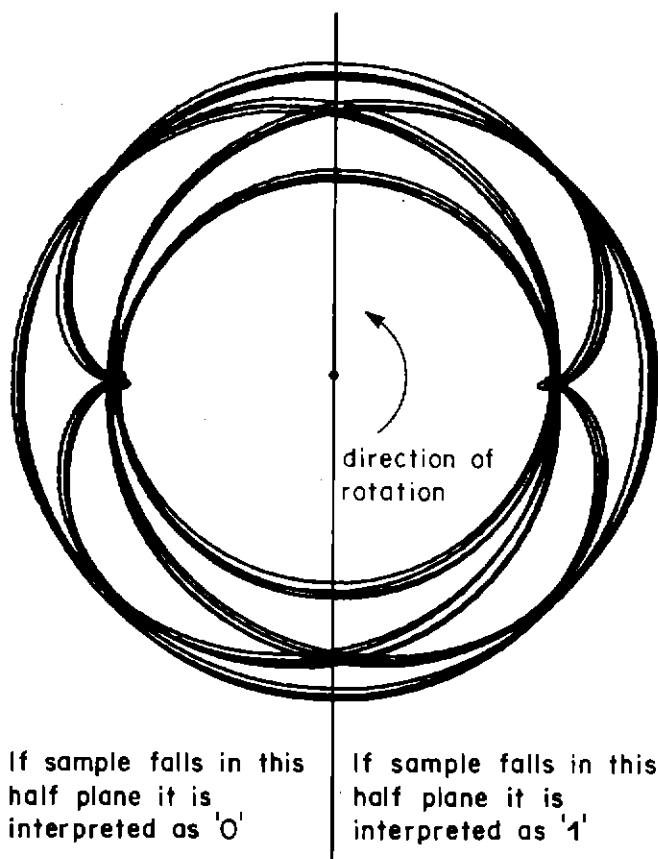


Fig. 9 - I.f. diagram showing decision threshold.

#### 4.2. Sensitivity to carrier phase perturbations and sample timing error

In the event that the local oscillator in the receiver has a phase error, there will be a degradation in performance. Phase errors can occur in the receiver by imperfect carrier recovery, or at any point in the transmission chain where phase perturbations can be induced into the signal.

When the signal is demodulated using the binary technique, compensation for phase errors can be obtained by modifying the sampling time instant.

Fig. 5(b) shows the i.f. diagram of a VSB 2-PSK signal. The amplitude is the instantaneous signal amplitude and the phase is the phase of the signal with respect to the carrier. If, when the signal is sampled the signal vector is situated in the right hand half of the plane, a logical '1' is assumed, and if it is in the left-hand half of the plane a logical '0' is assumed, as shown in Fig. 9.

If there is a carrier phase error the reference for the i.f. diagram is changed and so the diagram is rotated. As a consequence some of the sampling points move closer to the decision threshold and thus are more prone to errors in the presence of noise. Fig. 10 shows the position of the signal vector

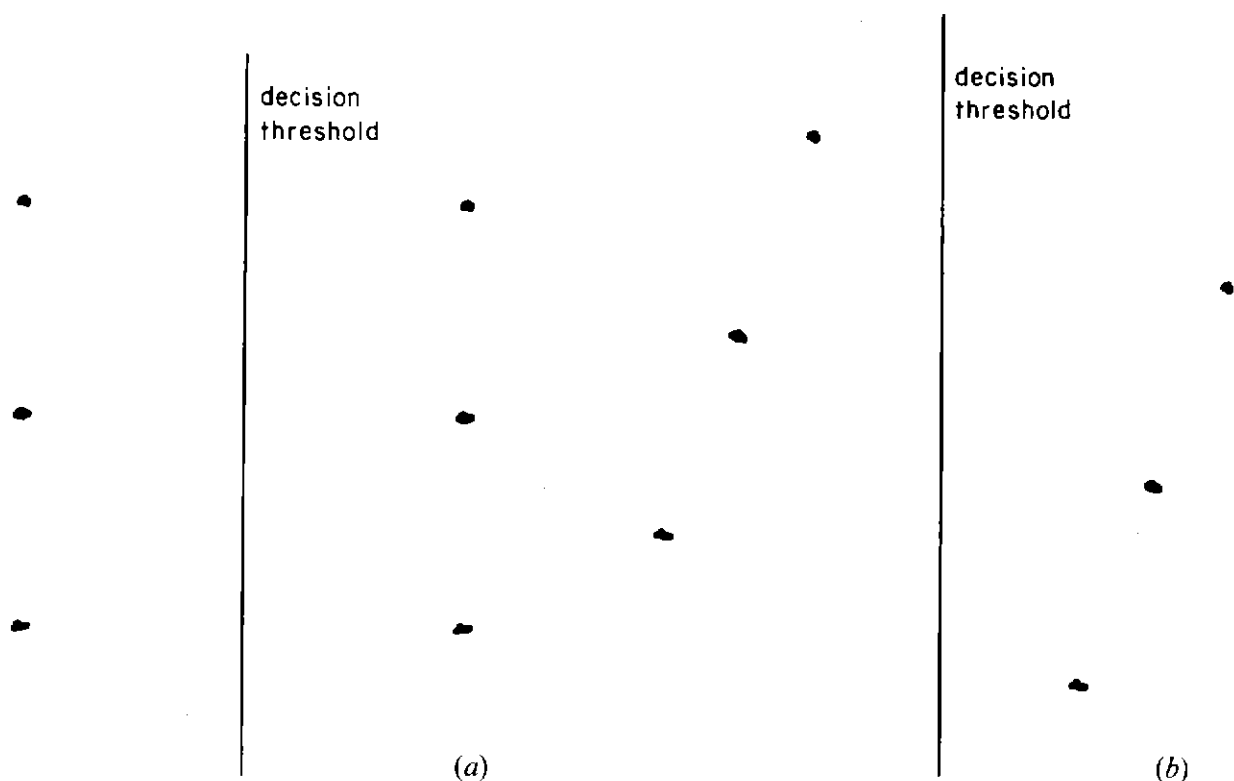


Fig. 10 - Sample diagram showing (a) correct phase of reference carrier, (b) incorrect phase of reference carrier: sample time as for (a) .

at the sampling instant

- a) with the correct reference carrier phase
- b) with an incorrect reference carrier phase but with the same sampling instant.

However, the signal vector always rotates in the same direction and so it is possible to compensate for phase errors by modifying the sampling instant. As can be seen from the eye diagram in Fig. 11, the

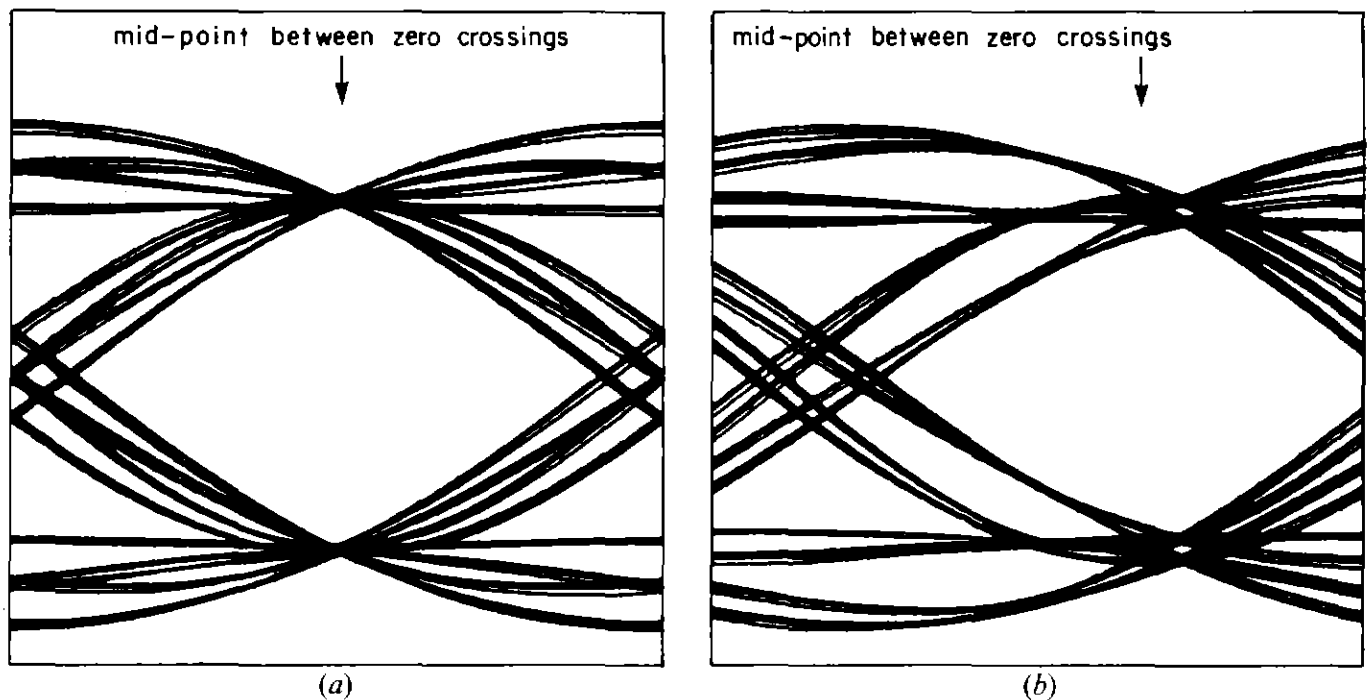


Fig. 11 - Eye diagram with, (a) correct phase of reference carrier, (b) incorrect phase of reference carrier.

optimum sampling time (i.e. the point of maximum eye opening) is very close to the sampling time derived by averaging zero-crossings. Thus the conventional method of timing recovery largely compensates for any errors in reference carrier phase. Fig. 12 shows the position of the signal vector with an incorrect reference carrier phase and sampling time derived from zero crossings, comparison with Fig. 10(b) shows the improvements.

Fig. 13 shows the way phase errors affect the error rate. In the absence of any timing compensation, errors increase rapidly as the reference phase departs from optimum. However, it is not realistic to consider demodulators which have no timing compensation: such a situation can only be achieved by using a 'cheat' wire from the modulator. When timing is taken from the zero-crossings of the demodulated waveform there is a noticeable improvement in error performance, and a small further improvement can be achieved if the point of max-

imum eye-opening can be identified. This can be accomplished by oversampling of the waveform: the technique used to assess the error performance by computer.

It is important to note that phase errors may be static (caused by faulty receivers for example) or dynamic (caused by picture related distortions). The performance of the receiver will depend on the time constants in the carrier and the clock recovery

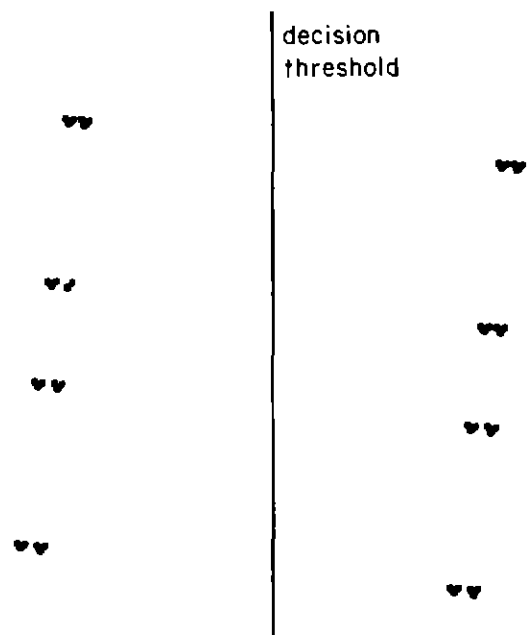


Fig. 12 - Sample time diagram with incorrect phase of reference carrier but with sample times modified to fall half-way between zero-crossings.

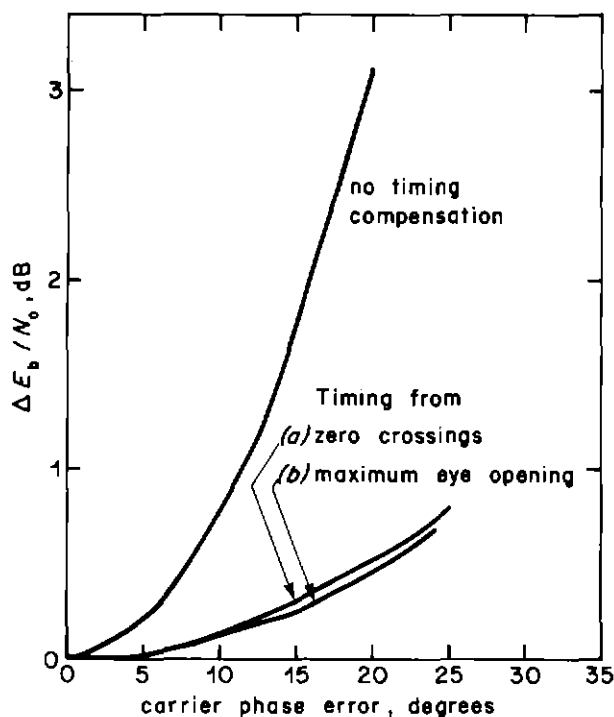


Fig. 13 - Change of  $E_b/N_0$  required to compensate for carrier phase error (computed).

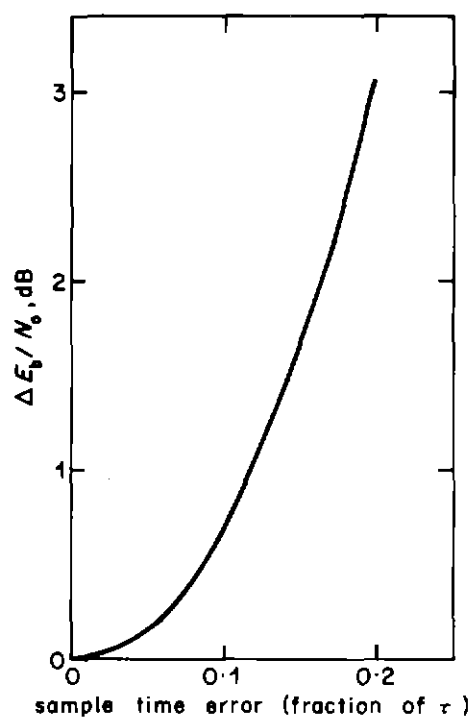


Fig. 14 - Change of  $E_b/N_0$  required to compensate for errors in sample time.

circuits. It is likely that full advantage can be taken of the properties of VSB 2-PSK when static errors exist. However, in the presence of dynamic errors the system may respond as if there were no timing compensation.

Fig. 14 shows the sensitivity of the system to errors in the sample time used in the receiver (compared with the optimum). VSB 2-PSK relies on the choice of optimum sample time to reduce the degradation caused by some distortions. It may be thought that as a consequence there would be great sensitivity to errors in sampling time. However, Fig. 14 shows that this is not the case. Typical errors in timing are small. Timing jitter is the main area of concern and typical values rarely exceed  $0.03 \tau$  r.m.s. The impairment from this will be scarcely measurable therefore.

#### 4.3. Sensitivity to channel imperfections

Any transmission channel may have imperfections in amplitude or group delay response. The imperfections will lead to a worse performance in the presence of noise than would otherwise have been the case.

The performance of VSB 2-PSK has been simulated on the computer to quantify the degradations in margin caused by channel imperfections.

The imperfections have been stylised; four types were used:

1. linear amplitude slope,
2. parabolic amplitude distortion,
3. linear group-delay slope,
4. parabolic group-delay distortion.

The amplitude variations are measured in dB and the group-delay variations in time across a band equal to half the bit rate. Note that linear amplitude slope and parabolic amplitude distortion are similar to the errors found when the centre frequency and receive bandwidth are inaccurate.

Figs. 15 to 18 show the reduction in noise levels needed to compensate for the introduction of distortion. (These curves are valid at a BER of  $10^{-3}$ ). Figs. 17 and 18 each show two curves corresponding to timing derived from the average of zero-crossings and to timing derived from maximum eye-opening. In the case of amplitude perturbations the carrier phase was kept constant whilst in the case of group delay perturbations it was found that the carrier phase could be optimised to maximise the eye-height.

These results show that VSB 2-PSK is not unduly sensitive to the types of distortion frequently found in transmission chains. In practice the most

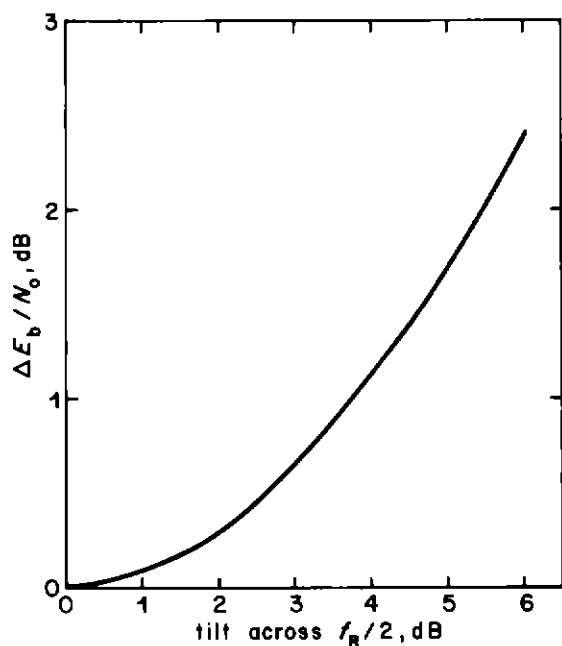


Fig. 15 - Change of  $E_b/N_0$  required to compensate for linear amplitude distortion.

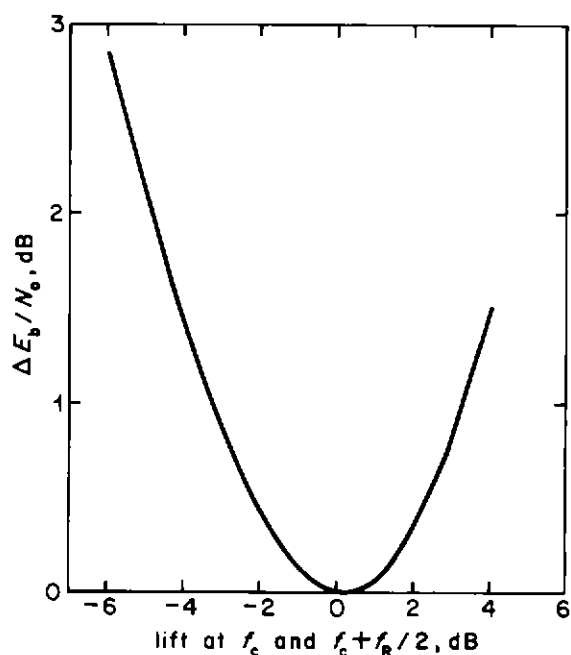


Fig. 16 - Change of  $E_b/N_0$  required to compensate for parabolic amplitude distortion.

likely distortion (especially in terrestrial channels) would be linear amplitude slope because the digital sound carrier is at the edge of the normal channel. Not only would there be a slope across the band but also severe attenuation of the whole signal. It is this overall attenuation which usually dominates the performance.

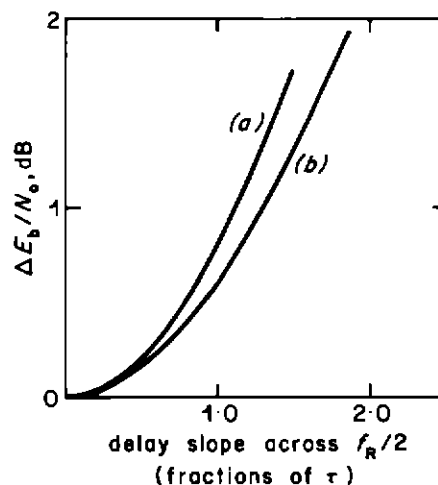


Fig. 17 - Change of  $E_b/N_0$  required to compensate for linear delay distortion,  
(a) sample time derived from zero-crossings,  
(b) sample time at maximum eye-opening.

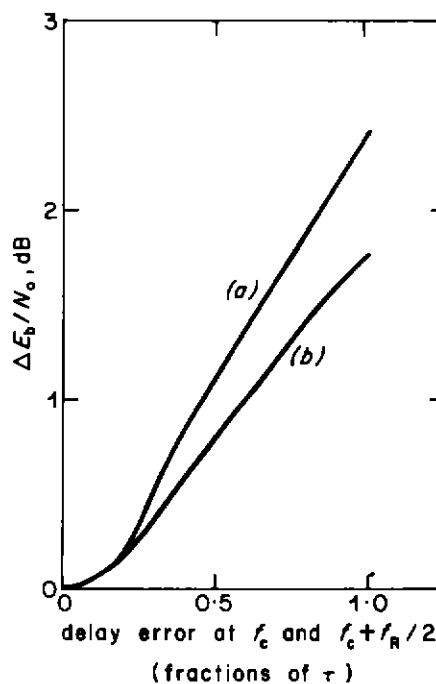


Fig. 18 - Change of  $E_b/N_0$  required to compensate for parabolic delay distortion,  
(a) sample time derived from zero-crossings,  
(b) sample time at maximum eye-opening.

## 5. Experimental work

### 5.1. Experimental configuration

Two experimental systems have been built, one operating at 2.048 Mbit/s for use in satellite studies and one for use at about 700 kbit/s for use in studies

of digital sound systems with terrestrial television (DSSTTV).

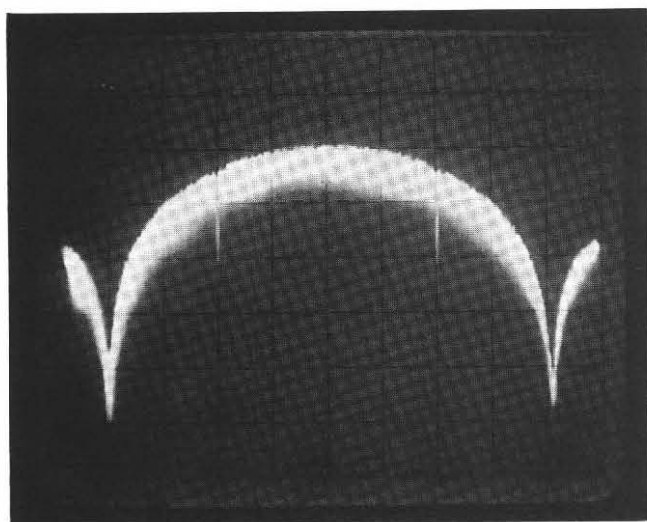
The first modulator was built using Pommier's variant<sup>10</sup> of MSK. This uses a simple resonator and delay line to convert 2-PSK into MSK. The MSK signal was then filtered using phase-equalised Butterworth and Gaussian filters to give the desired VSB 2-PSK signal.

The second modulator generated VSB 2-PSK directly using digital techniques. The signal was

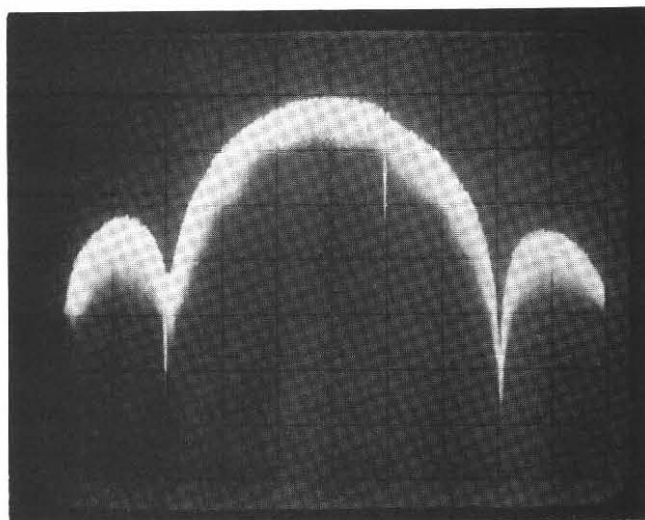
constructed from samples at four times the bit-rate frequency directly using a mathematical algorithm.

Fig. 19 shows the signal spectra. These spectra may be compared with the theoretical curves in Section 2.

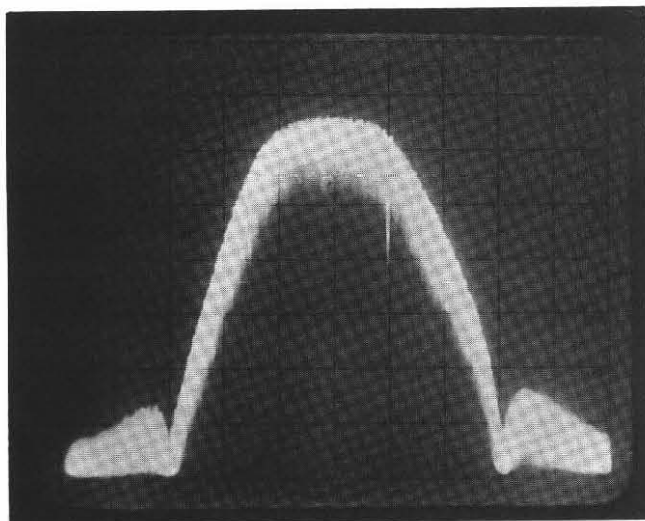
The receiver operating at 2.048 Mbit/s used frequency doubling to recover the reference carrier. After input filtering (using Gaussian shaping) the signal was frequency doubled. The strong components at twice the carrier frequency (see Fig. 20) were



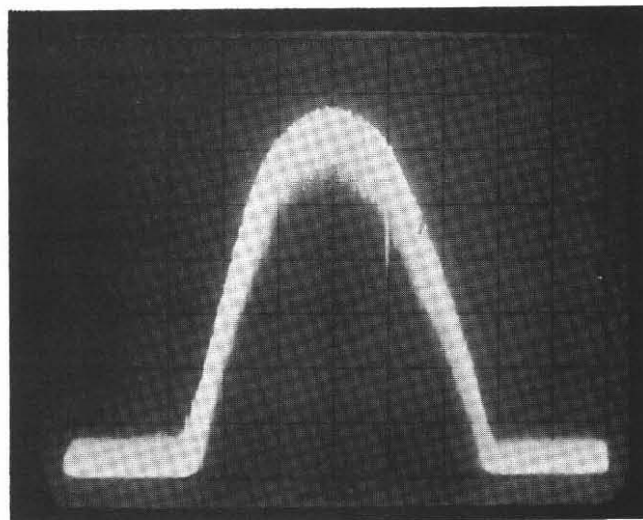
(a)



(b)



(c)



(d)

*Fig. 19 – Measured spectra,  
(a) 2-PSK,  
(b) MSK,  
(c) VSB 2-PSK on transmission,  
(d) VSB 2-PSK at the receiver.*

Vertical scale 10 dB/div.  
Horizontal scale 0.5 MHz/div.  
Bit-rate 2.048 Mbit/s.

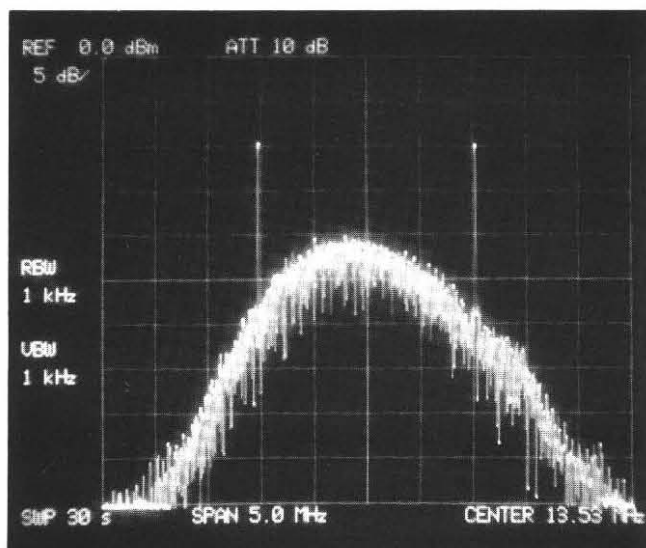


Fig. 20 – Spectrum of squared signal in carrier recovery circuit. The left-hand peak is at twice carrier frequency. That on the right is offset by the bit-rate.

Vertical scale 5 dB/div.  
Horizontal scale 0.5 MHz/div.

then recovered in a PLL. The recovered carrier was used to synchronously demodulate the modulated signal to give a baseband data signal. This was then sliced and sampled using clock timing recovered from the zero crossings of the demodulated signal.

The receiver operating at about 700 kbit/s used a modified Costas loop to recover the carrier. Although a quadrature process was used for carrier recovery, the demodulation used was a binary process.

The experimental results quoted here are for the 2.048 Mbit/s system. Those for the lower bit-rate are similar, though less comprehensive measurements have been made.

## 5.2. Performance in the presence of noise

The effect of noise on the error rate is shown in Fig. 21. The noise was added to the signal in the transmission path. It can be seen that the performance is close to that expected from the theory outlined in Section 4.1.1.

## 5.3. Effect of carrier phase errors

The effect of errors in the phase of the recovered carrier is shown in Fig. 22. The phase of the carrier could be modified by inserting delay in between the carrier recovery circuit and the demodulator. The clock timings are recovered using the zero crossings of the signal as a reference. Fig. 23 shows the effect of carrier phase errors on the received eye. The syn-

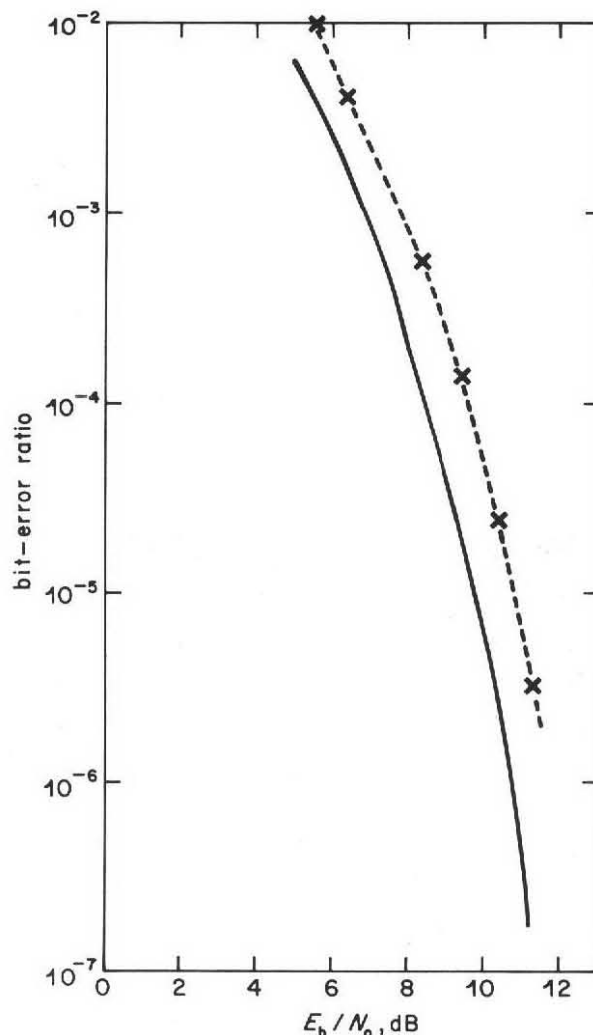


Fig. 21 – Measured relationship between error rate and noise levels,

— theory,  
--- measurement.

ronisation is derived from the clock in the modulator. It can be seen that the optimum sampling time has shifted as predicted in Section 4.2.

## 5.4. Effect of multipath propagation

The effect of multipath propagation was assessed by using a technique similar to that used by Kallaway<sup>11</sup> for QPSK. A delayed signal was attenuated and added to the main signal. Noise was then added to the two signals representing direct and indirect waves, and the resultant signal demodulated (See Fig. 24).

With infinite attenuation in the echo signal path, noise was added to give an error rate of 1 in  $10^3$ . The noise level was then attenuated by 1 dB and the level of the delayed signal increased to give the same error rate. This was repeated for a range of

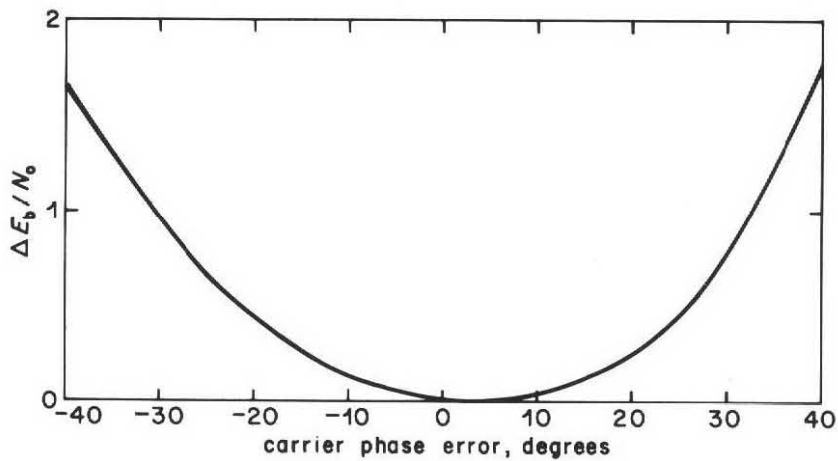
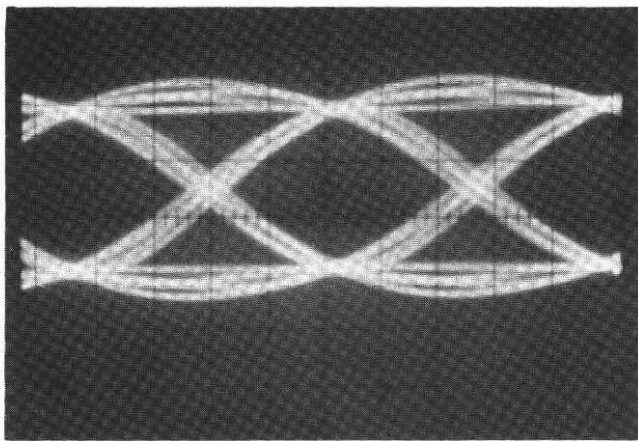
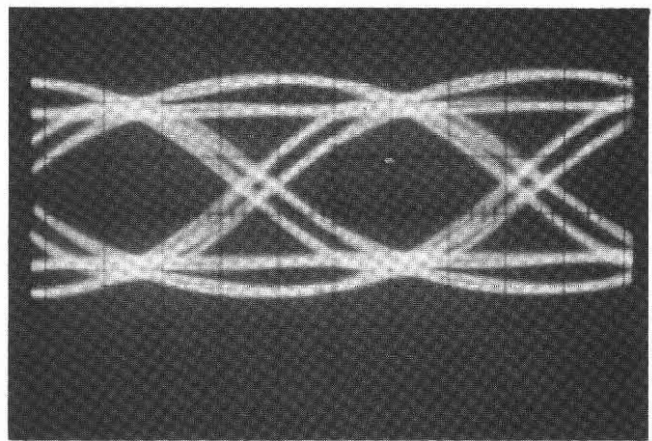


Fig. 22 – Measured change of  $E_b/N_0$  required to compensate for carrier phase errors.



(a)



(b)

Fig. 23 – Measured eye patterns,  
(a) No carrier phase error,  
(b) Carrier phase error deliberately introduced.

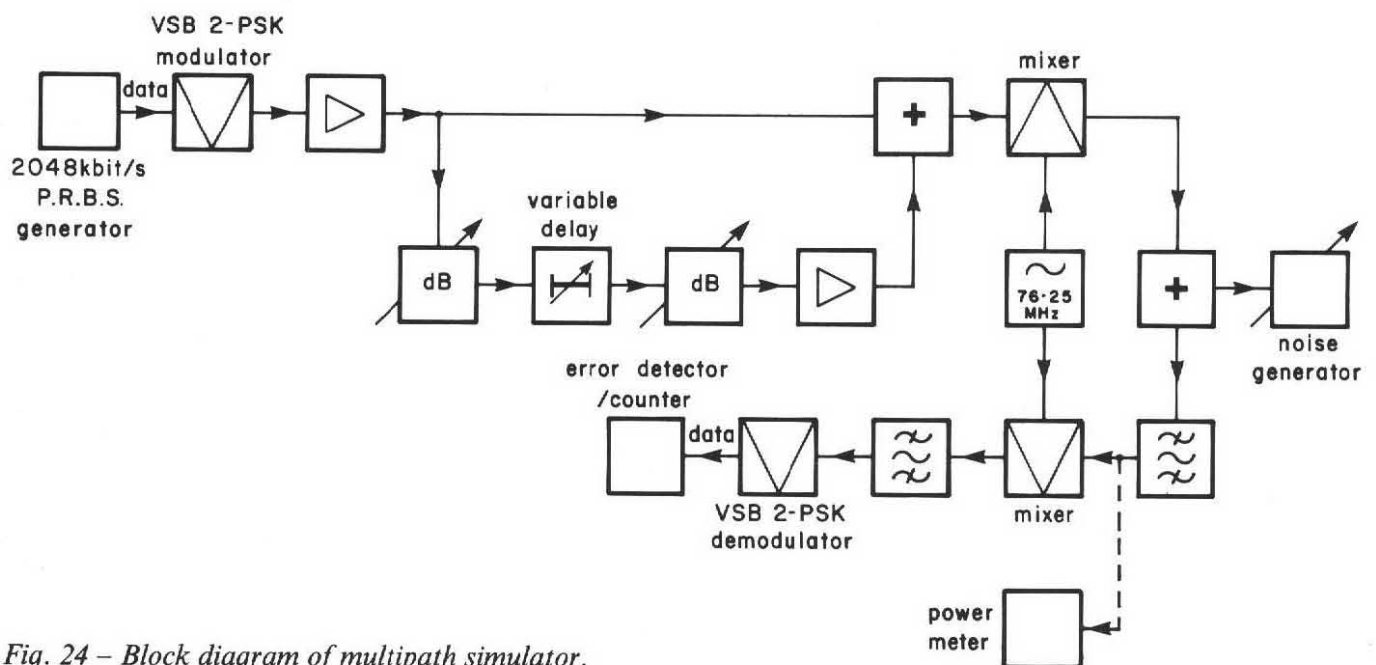


Fig. 24 – Block diagram of multipath simulator.

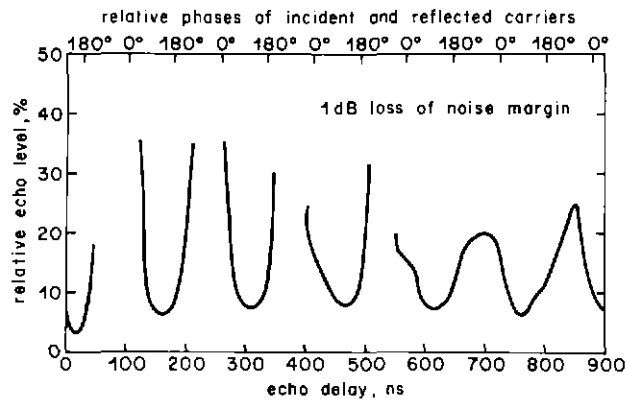


Fig. 25 - Level of short echo needed for 1 dB loss of noise margin.

delays from a few degrees at carrier frequency to over 4 data bit periods. The experiment was repeated with the noise attenuated by 2 dB and 3 dB from that necessary to give a BER of  $10^{-3}$ .

Fig. 25 shows the susceptibility of the system to echoes for a range of delays and constant 1 dB loss of noise margin. At delays of less than 1 bit period (about 488ns) the points of minimum susceptibility are at infinite echo level and occur when the direct and reflected carriers are in phase; this enhances the incident wave (as the incident and reflected waves are indistinguishable), thus improving the performance of the system. Maximum susceptibility occurs when the incident and reflected waves were measured to be approximately in antiphase. Above 488ns however, maximum susceptibility occurs when the signals are near quadrature. Fig. 26 shows the effect of longer term delays at 1 to 3 dB loss of noise margin. At delays above one bit period the difference between best and worst case (due to carrier phasing) is less than that observed below 1 bit period delay.

Fig. 26 compares directly with Kallaway's Fig. 4a, b and c in Reference 11. It can be seen that for a given loss of noise margin, the mean level of interference is the same for QPSK and VSB 2-PSK. From this it appears that VSB 2-PSK is no more vulnerable to multipath propagation than QPSK. However recent results<sup>12</sup> show that QPSK does perform better than VSB 2-PSK over a small range of carrier phase differences when the echo delay is between  $\frac{1}{2}$  and  $1\frac{1}{2}$  bit periods.

## 6. Conclusions

Vestigial sideband binary phase-shift keying (VSB 2-PSK) is a bandwidth efficient variant of 2-PSK. Experimental measurements have confirmed

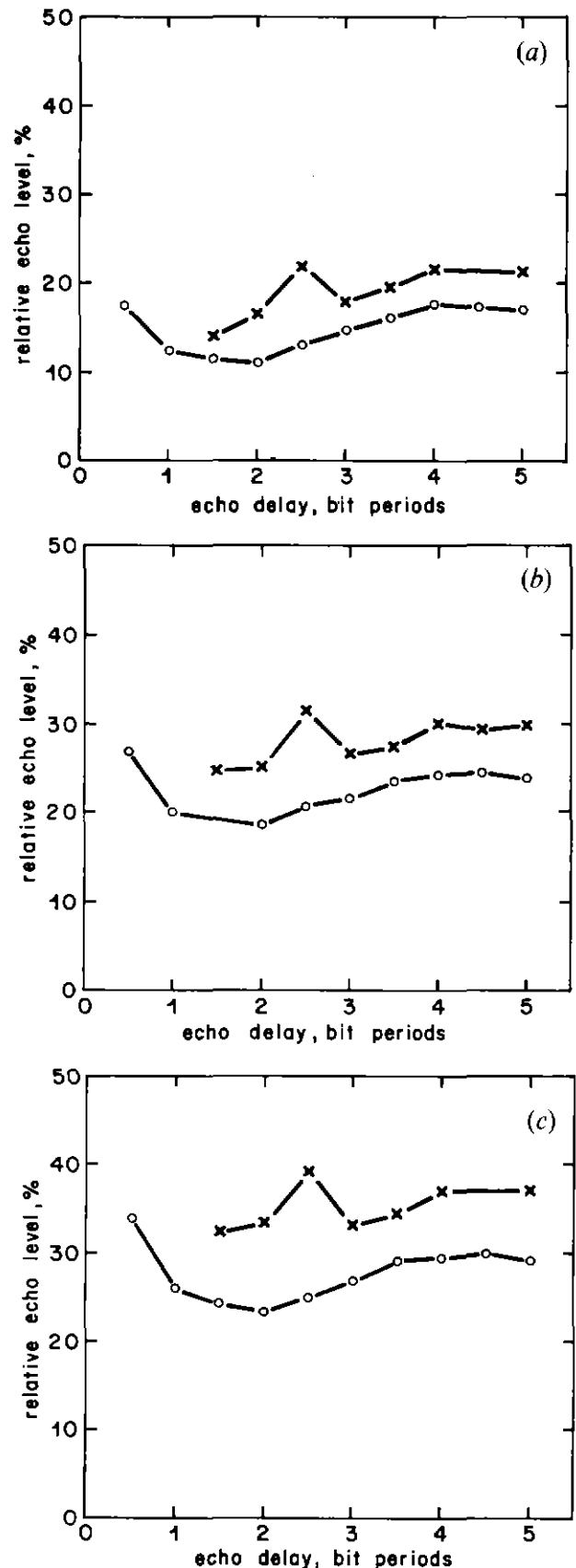


Fig. 26 - Level of echo required for, (a) 1 dB, (b) 2 dB and (c) 3 dB loss of margin,   
 ○—○ worst phase,   
 ×—× optimum phase.

that a performance close to theoretical can be achieved.

A feature of VSB 2-PSK is that it may be received with a binary demodulator and simple filtering which makes it potentially attractive for the production of cheap receivers. It is not unduly sensitive to the distortions typically found in a transmission channel and this has been verified experimentally in particular cases for a digital signal of modest capacity carried above a vision signal. Thus VSB 2-PSK is a suitable modulation system for use on a carrier associated with a television waveform.

VSB 2-PSK was chosen by the EBU as the most suitable modulation system if a subcarrier with f.m. is used for satellite television broadcasting. Although also suitable for use on a carrier with existing terrestrial television transmissions, QPSK has been chosen for the U.K. 2-channel digital sound with television system, because of availability of integrated circuits and small differences in the relative importance of the various performance factors in a terrestrial channel.

## 7. Acknowledgements

Much of the work reported here was inspired by discussions within the EBU. The authors would like to thank Mr. D. Pommier of CCETT and Dr. R.A. Harris of ESA for their development of the system and all the other members of the EBU whose comments and suggestions have helped in our understanding. The authors would also like to thank Mr. Pommier and his colleagues at CCETT for the loan of modulation equipment and their support and advice during the early experiments.

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## Appendix 1

**Relationship between signal-to-noise ratio in the vision channel and  $E_b/N_o$  of the data signal for a terrestrial system.**

In a terrestrial system the noise spectrum is substantially flat before demodulation. After demodulation the noise is flat in regions not affected by the vestige filtering.

Let  $N_o$  (W/Hz) be the noise spectral density before i.f. filtering and demodulation

Let  $X^2$  be the mean power in the digital signal

Let  $Y^2$  be the mean power of the vision signal at the peak of the sync. pulse.

Thus

$$E_b/N_o = X^2/RN_o. \quad (A-1)$$

The video carrier-to-noise ratio  $C/N$  is given by

$$C/N = 10 \log_{10} Y^2/N_o B \text{ (dB)} \quad (A-2)$$

where  $B$  is the effective i.f. noise bandwidth.

Also the signal-to-noise ratio after demodulation is related<sup>13</sup> to the carrier-to-noise ratio before demodulation by\*

$$S/N = C/N - 8 \text{ (dB)} \quad (A-3)$$

where the video signal-to-noise ratio is measured in a 5.5 MHz bandwidth and is not weighted. The effective i.f. noise bandwidth ( $B$ ) is 5.08 MHz.

Therefore

$$S/N = 10 \log_{10} (Y^2/N_o B) - 8 \text{ (dB)} \quad (A-4)$$

$$S/N = E_b/N_o + 10 \log_{10} Y^2/X^2 - 10 \log_{10} B/R - 8 \text{ (dB)} \quad (A-5)$$

$$E_b/N_o = S/N - 10 \log_{10} Y^2/X^2 + 10 \log_{10} B/R + 8 \text{ (dB)} \quad (A-6)$$

This relationship shows how the level of the digital carrier depends on the required video signal-to-noise ratio and digital  $E_b/N_o$  at failure. It does not include any factors to allow for signal degradation. In practice the  $E_b/N_o$  will have to be larger by a margin which takes into account distortion, imperfect equipment, multipath propagation and a safety margin.

\* This is not the same relationship as that quoted by Ely.<sup>8</sup> He bases his result on Osborne who uses a reference corresponding to a mean power level rather than peak sync, and an i.f. noise bandwidth of 6.25 MHz.

For example, if the digital signal is 20 dB below the peak power of the vision, the vision bandwidth used is 5.5 MHz, and the bit rate is 728 kbit/sec, then

$$E_b/N_o = S/N - 3.6 \text{ (dB)} \quad (A-7)$$

## Appendix 2

**Relationship between carrier-to-noise ratio, vision signal-to-noise ratio and  $E_b/N_o$  for a f.m. channel**

In a f.m. channel above threshold the noise spectrum before demodulation is substantially flat. After demodulation it is triangular (in voltage terms) or quadratic (in power terms).

The conventional way to derive the video signal-to-noise ratio from the carrier-to-noise ratio assumes a noise power density,  $\eta$ . Noise in a band  $\delta f_c$  offset  $f$  from the carrier at  $f_c$  produces noise with baseband frequency  $f$  and a power  $P(f)$ ,

$$P(f) = \eta f^2 \delta f / A_c^2 \quad (A-1)$$

where  $A_c$  is the carrier amplitude. It is assumed that the demodulator has unit sensitivity. Equation A-1 is a well known expression and when integrated over the range of video frequencies which concern us, 0 to  $f_v$ , leads to the usual relationship between video  $S/N$  and  $C/N$ .

$$\text{Noise power} = \int_{-f_v}^{f_v} \frac{\eta f^2 df}{A_c^2} \quad (A-2a)$$

$$= \frac{2}{3} \cdot \frac{\eta f_v^3}{A_c^2} \quad (A-2b)$$

$$\text{but } C/N = \frac{A_c^2}{2} \cdot \frac{1}{\eta B} \quad (A-3)$$

where  $B$  is the i.f. bandwidth, and

$$\text{Signal power} = K^2 \Delta f^2 \quad (A-4)$$

where  $\Delta f$  is the carrier deviation produced by a 1V video signal and  $K$  is the reference video amplitude.

$$S/N = \frac{K^2 \Delta f^2}{\frac{2}{3} \eta f_v^3 / A_c^2} \quad (A-5a)$$

$$= \frac{3BK^2 \Delta f^2}{f_v^3} \cdot \frac{C}{N} \quad (A-5b)$$

with PAL,  $K$  is usually taken to be 0.7V. Equation (A-5b) then reduces to

$$S/N = \frac{3}{2} \cdot \frac{B \Delta f^2}{f_v^3} \cdot \frac{C}{N} \quad (A-6)$$

For example the usual values assumed for DBS are

$$\begin{aligned} B &= 27 \text{ MHz} \\ \Delta f &= 13.5 \text{ MHz pk-pk} \\ f_v &= 5.5 \text{ MHz} \end{aligned}$$

in this case

$$S/N = C/N + 16.5 \text{ (dB)} \quad (\text{A-7})$$

This equation does not assume pre-emphasis or weighting.

Pre-emphasis according to CCIR Rec.405 produces

$$S/N = C/N + 18.5 \text{ (dB)} \quad (\text{A-8})$$

and with pre-emphasis and weighting (CCIR Rec 410.2) the equation becomes

$$S/N = C/N + 30.3 \text{ (dB)} \quad (\text{A-9})$$

The use of  $E_b/N_o$  in modulation theory normally assumes that the noise spectrum is flat. In this instance it is not, and so all the results obtained are not strictly valid. However, in practice the Gaussian properties of the noise are retained and the differences between analysis based on flat noise and that based on triangular noise are very small. Thus it is possible to make the assumption that the noise level is flat with a spectral density equal to that found at mid-band.

Thus if  $f_o$  is the centre frequency

$$N_o = 2\eta f_o^2 / A_c^2 \quad (\text{A-10})$$

Again assuming the demodulator has unit sensitiv-

ity, the r.m.s. deviation of the main carrier produced by the subcarrier is  $\Delta f$  and  $R$  is the bit rate,

$$E_b = \Delta f^2 / R \quad (\text{A-11})$$

$$\begin{aligned} E_b/N_o &= \Delta f^2 / R \cdot A_c^2 / 2\eta f_o^2 \\ &= \Delta f^2 / R \cdot B / f_o^2 \cdot C/N \end{aligned} \quad (\text{A-12})$$

If the r.f. bandwidth is 27 MHz, equation (A-12) can be rewritten

$$\begin{aligned} E_b/N_o &= C/N + 20 \log_{10} \Delta f / 2.5 - 20 \log_{10} f_o / 7 \\ &\quad - 10 \log_{10} R / 2 + 2.3 \text{ (dB)} \end{aligned} \quad (\text{A-13})$$

where all the frequencies and bit-rates are in MHz and Mbit/sec.

If  $\Delta f = 2.5 \text{ MHz}$ ,  $f_o = 7 \text{ MHz}$  and  $R = 2 \text{ Mbit/sec}$

$$E_b/N_o = C/N + 2.3 \text{ (dB)}$$

If vision failure is taken to occur at threshold ( $C/N = 10 \text{ dB}$ ) and sound failure at a BER of  $10^{-3}$  (corresponding to  $E_b/N_o$  of 6.8 dB) then the digital carrier has a margin of 5.5 dB.

These equations are valid even beyond threshold because in the region of the subcarrier the triangular noise dominates the flat noise spectrum of the threshold spikes. Threshold spikes have to be considered for very low values of  $C/N$  (less than 4 dB say). Care has to be taken however, to ensure that no problems are caused by non-linearity in the video circuits following the demodulator. Non-linearities here can mean that the error performance is directly related to the number of threshold spikes, even at  $C/N$  well above those normally associated with threshold.